Wide Band CMOS Low-Noise Amplifier Exploiting Noise Cancellation

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Abstract—**Well-known elementary wide band amplifiers suffer from a fundamental trade-off between noise factor and source impedance matching, which limits their noise figure (NF) to values typically above 3dB. Negative feedback can be employed to break this trade-off, thus allowing lower noise figures, however, at the price of potential instability. In contrast, this paper describes a novel feed-forward noise cancelling technique, which allows for design wide band Low Noise Ampliers (LNAs) with sub 3dB noise figure, without suffering from instability issues. This technique is also robust and easy to use. Measurements on a wide band LNA realised in 0.25μm standard CMOS show NF values below 2.4dB over more than one decade of bandwidth and below 2dB over more than two octaves.**

Keywords— **low-noise amplifier, LNA, wide band, broad band, noise cancelling, noise cancellation.**

I. INTRODUCTION

Wide band Low-Noise Amplifiers (LNAs) constructed using MOSTs and resistors are traditionally found in receiving systems where the ratio between the signal bandwidth (BW) and its center frequency can be as large as one. Good examples are analog cable (50-850MHz), digital satellite (950-2150MHz) and digital terrestrial (450-850MHz) video broadcasting. At these frequencies, parallel and series LC tanks used as load and for source degeneration do not perform adequately over the bandwidth or integrated coils are impractical due to their large size (i.e. large area, modest Q and low self-resonance). In alternative, a wide band inductorless LNA can replace parallel LC tuned LNAs in multi-band and multi-mode receivers. In this case, a wide-band LNA uses much less chip-area and increases the flexibility of the front-end. Furthermore, when more frequency bands need to be received simultaneously, the typically larger power dissipation of a wide band LNA is not an issue as the total power of the tuned LNAs is available. This last point does not hold for a concurrent multi-band LNA [1] because only one active device is combined to matching networks with multiple resonance. However, in low-cost MOS processes, the relatively low quality of inductors limits the possibility to accommodate continuous bands.

In general, high-sensitivity highly-integrated receivers require high-performance LNAs with sufficiently large gain, noise figures well below 3dB, adequate linearity and source impedance matching $Z_{IN}=R_S$. The latter is wanted in order to avoid cable reflections or alterations of the characteristics of the RF filter preceding the LNA [2], such as pass-band ripple and poorer stop-band attenuation. This combination of requirements must be guaranteed over a wide band of frequencies, while also allowing for some variable gain to handle interference generated by strong adjacent channels.

Traditional wide band LNAs built of MOSTs and resistors have difficulties in meeting the combination of previously mentioned requirements. Especially, many elementary amplifiers [3,4,5] fail to achieve low noise figures upon $Z_{IN}=R_S$. Amplifiers based on the use of negative feedback can achieve low noise but are prone to instability [6]. In this paper, a novel wide-band noise cancellation technique is presented, which can achieve very low noise figure without instability problems. The technique was validated through the design of a sub-2dB noise figure wide-band LNA in a 0.25μm CMOS [7].

The noise cancellation technique presented in this paper was devised using voltage controlled current sources as building blocks [4,12]. MOS transistors operating in strong inversion and saturation approximate this behaviour for small signals. Unless otherwise stated, a MOS transistors will be regarded as a voltage controlled current source, $I=g_m \cdot V$, with *zero* output conductance. Channel thermal noise is assumed the dominant source of noise. This is a realistic assumption for a properly designed wide band LNA of bandwidth $[f_a, f_b]$ with $f_a >> f_{1/f}$ and $f_b << f_T$, where $f_{1/f}$ is the 1/f noise corner frequency and $f_T = g_m/(2\pi C_{gs})$ the unity current gain cut-off frequency.

The paper is organised as follows. Section II reviews properties and basic limitations of known monolithic wide band LNAs. Section III introduces the novel noise cancelling technique. Section IV describes the design of a wide band LNA in 0.25µm standard CMOS. Section V deals the measurement results of the demo chip. Finally, section VI draws the conclusions.

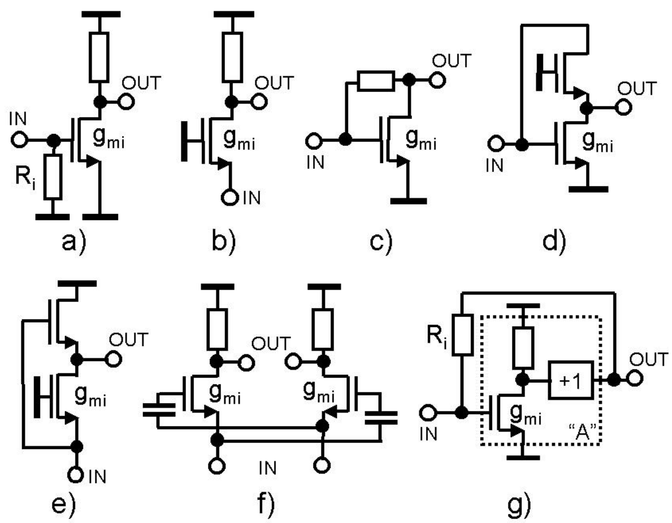


Figure 1: Known wide band LNAs (biasing not shown).

II. REVIEW OF EXISTING TECHNIQUES

In this section, commonly used wide band low-noise techniques are reviewed in order to highlight their limitations, thus placing in perspective the new proposed technique. To this purpose, the power spectral density of the channel noise is expressed as $4kT \cdot NEF \cdot g_m \cdot \Delta f$, with $NEF = \gamma \cdot (g_{d0}/g_m)$ the noise excess factor of the device and γ the well-known bias-dependent noise constant. For a deep sub-micron MOS, $(g_{d0}/g_m)>1$ and $1<\gamma<2$ hold [8]. This modelling can be used also for a resistor if $g_m=1/R$ and NEF=1 are used.

A. NF to $Z_{IN}=R_S$ trade-off in elementary amplifiers

Figure 1a-e shows well-known elementary wide band amplifiers: the resistively terminated common-source stage (a), the common-gate stage (b), the common-source stage with resistive shunt-feedback (c) and the circuits found in [3,4] (d and e). Their input impedance is $Z_{IN} \in \{1/g_{m,i}, R_i\}$ and $g_{m,i}$ is the transconductance of the input MOST. These amplifiers suffer from a severe trade-off between their noise factor F (NF=10*log(F)) and the matching requirement, $Z_{IN}=R_S$. For sufficiently large LNA gain, the input device dominates the noise and low values of F require an input device $g_{m,i}R_S$ and R_i/R_S much larger than one. In contrast, $Z_{IN}=R_S$ demands a fixed $g_{m,i}R_S=1$ and $R_i/R_S=1$. Then, for $Z_{IN}=R_S$, F is above $1+NEF \geq 2$ (i.e. NF>3dB) because of the matching device contribution (NEF).

The trade-off between F and $Z_{IN}=R_S$ is somewhat relaxed for a balanced common-gate amplifier exploiting capacitive input cross-coupling [5] (figure 1f). Due to the cross coupling, upon a differential impedance match, the g_{mi} is two times smaller. The contribution to F of the two input MOSTs is then equal to NEF/2, which is two times smaller than for the traditional balanced common gate stage. This means that, for equal gain, the use of cross coupling provides a lower F while requiring half of the power! Nevertheless, cross coupling suffers from some important limitations:

- F cannot be better than 1+NEF/2 because the trade-off with $Z_{IN}=R_S$ stands still. For noise demanding applications, 1+NEF/2 is just not enough.
- Antennas, cables and cheap high-frequency filters are usually single-ended devices. In this case, a balun in front to the LNA can be used to perform the single-ended to balanced conversion. However, a wide-band low-loss discrete balun is relatively expensive, occupies PCB area and couples extra interference at the input node. Also, the balun loss increases the cascade noise factor $F_{Balun+LNA}$.
- Voltage drop across the cross coupling capacitances degrades F, unless Cs are large (e.g.: 2x10pF [5]).

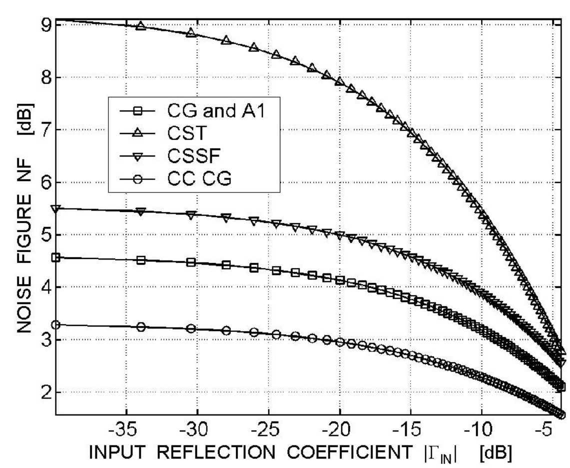


Figure 2: LNA noise figure vs the input reflection coefficient for equal power (A_{VF}=-10, NEF=1.5).

The trade-off between F and $Z_{IN}=R_S$ can be relaxed by deliberately mismatching the source in such a way that a lower F is achievable. This is possible because accurate matching is seldom required while an input reflection coefficient $|\Gamma_{IN}=(Z_{IN}-R_S)/(Z_{IN}+R_S)|$ between -8 and -10 dB is sufficient in most applications. Figure 2 shows the noise figure of the amplifiers in figure 1a-f for $g_{m,i}R_S>1$ ($\Gamma_{IN}<0$) or $R_i/R_S>1$ ($-\Gamma_{IN}>0$), assuming equal voltage gain $|A_{VF,TOT}=V_{OUT}/V_S|=5$ and NEF=1.5. Clearly, for $|\Gamma_{IN}|_{dB}$ close to -10dB, only the cross-coupled amplifier provide NF values below 3dB, until NF reaches its minimum for $|\Gamma_{IN}|_{dB}=|\Gamma_{IN}|_{dB,MAX}$.

B. Breaking the trade-off via negative feedback

The trade-off between F and $Z_{IN}=R_S$ is broken when the (equivalent) input noise of the matching device can be lower than the noise of the source. This can be performed exploiting negative feedback. In theory, non-energetic transformer-feedback amplifiers provide superior performance [8,9,10]. In practice, because of the poor quality of integrated wide-band transformers, resistive feedback is most practical. Figure 1g shows perhaps the simplest example of a single-loop wide-band amplifier: a voltage amplifier "A" with a shunt-feedback resistor R_i between its input and output terminals. Its input impedance is $Z_{IN}=R_i/(1-Av)$ and Av is its *negative* gain of amplifier "A".

To understand how this amplifier can provide a low F upon a matched input lets initially assume that Av is zero (i.e. loop open). In this case, R_i acts as a resistive source termination and for $R_i=R_S$ the noise factor $F=1+I^2_{n,Ri}/I^2_{n,RS}=1+R_S/R_i$ is 2 (i.e. NF=3dB). When Av is non-zero (i.e. loop active), the noise current of R_i adds still to the noise of the source because of the zero output resistance of amplifier "A". However, for $Z_{IN}=R_S$, R_i is now (1-Av) times *larger* than R_S so its noise current is (1-Av) times *smaller*. Therefore, the noise factor $F=1+I^2_{n,Ri}/I^2_{n,RS}=1/(1-Av)$ is small, provided adequate gain Av is available. In reality, amplifier "A" generates some noise on its own, thereby degrading F. However, "A" is not constrained by matching, so its contribution to F can be arbitrarily small by increasing the g_mR_S of its input stage. Therefore, for a proper gain Av, the feedback amplifier in figure 1g is capable of noise figure values well below 3dB provided the contribution of amplifier "A" is minimised. Despite its properties amplifiers based on figure 1g suffer from some drawbacks:

- Sufficient gain and GHz bandwidth often mandates the use of multiple-cascaded stages *within* the feedback loop (two in figure 1g), thereby making its operation prone to instability.
- Z_{IN} depends on all the loop parameters, so it is rather sensitive to process-spread.
- Since Z_{IN} and Av are directly coupled, variable gain at $Z_{IN}=R_S$ is not straightforward.
- For $Z_{IN}=R_S$, the loop gain 1+T=1-Av/(2-Av) is ever lower than 2 (i.e. most of the feedback voltage is across R_i). Therefore, the closed-loop linearity is NOT significantly better than that of amplifier "A"[12]. Even more, when "A" consist of cascaded stages and most of the gain is allocated in the first one (i.e. for noise reasons), linearity can be poor [6].

III. THE NOISE CANCELLING TECHNIQUE

In this section, a novel wide band low noise technique is presented, which is able to de-couple F from $Z_{IN}=R_S$ without global negative feedback or compromising the source match. This is achieved by *cancelling* the output noise contribution of the matching device *without* degrading signal transfer. After all, if the matching device contributes no output noise, it does not affect F, so the trade-off is broken.

To understand how this noise cancelling works let's begin with a well-known wide-band matched amplifier: the common source stage with shunt-feedback (figure 3). This amplifier has an input impedance $Z_{IN}=1/g_{mi}$, gain $A_{VF}=V_Y/V_X=1-g_{mi}R$ and output resistance $R_{OUT}=(R+R_S)/(1+g_{mi}R_S)$, which make it suitable for high frequency applications. However, as previously mentioned, its F for $Z_{IN}=R_S$ is larger than 1+NEF due to the contribution of the matching device.

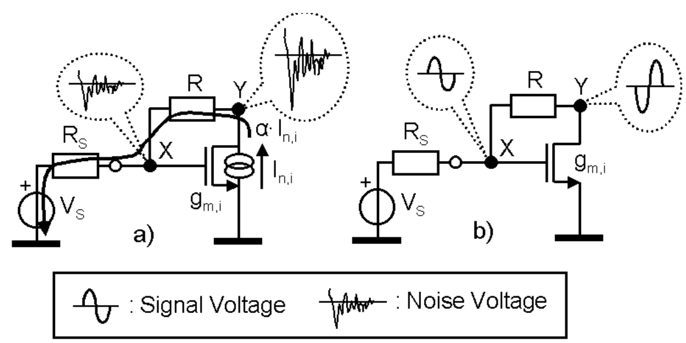


Figure 3: Matching device noise (a) and signal (b) at nodes X and Y

Let's now focus on the signal and the matching device noise voltages at the input node X and output node Y. Figure 3a sketches the instantaneous noise voltages for a given direction of the matching device noise current. Depending on the relation between $Z_{IN}=1/g_{m,i}$ and R_S, the noise current $\alpha(R_S,g_{m,i})\cdot I_{n,i}$, ($\alpha$ is a constant between 0 and 1), flows out the matching device. For instance, for $Z_{IN}=R_S$, α is ½ so only half the channel current flows out from the matching device while the rest flows inside it. Since R and R_S are connected in series, the noise current

$\alpha(R_S,g_{m,i})\cdot I_{n,i}$ generates two instantaneous noise voltages at nodes X and Y that have different amplitudes but *equal sign*. On the other hand, figure 3b shows that the signal voltage at nodes X and Y to ground have different amplitudes but *opposite* sign because the gain A_{VF} is negative for $g_{mi}R>1$. This sign-difference between the instantaneous noise and signal voltage at nodes X and Y enables the possibility to cancel the noise of the matching device while *simultaneously* adding signals. This is done by delivering to a new output the sum of the voltage at node Y plus the voltage at node X multiplied by a proper *negative* scaling factor, which equalises both the amplitude and the sign differences. Figure 4 shows the implementation of noise cancelling using an ideal feed-forward voltage amplifier "A" in parallel to the matching stage.

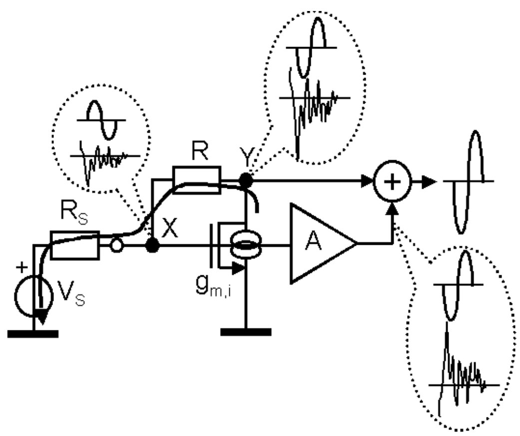


Figure 4: Wide band LNA exploiting noise cancelling.

By circuit inspection, the matching device noise voltage at node X, ($V_{X,n,i}$) and Y ($V_{Y,n,i}$) are equal to:

$$V_{X,n,i}=\alpha(R_S,g_{m,i})\cdot I_{n,i}\cdot R_S$$
$$V_{Y,n,i}=\alpha(R_S,g_{m,i})\cdot I_{n,i}\cdot(R+R_S)$$

The output noise voltage, $V_{OUT,n,i}$ is then equal to:

$$V_{OUT,n,i}=V_{X,n,i}\cdot Av+V_{Y,n,i}=\alpha(R_S,g_{m,i})\cdot I_{n,i}\cdot(R+R_S+Av\cdot R_S)$$

Output noise cancellation $V_{OUT,n,i}=0$ is then achieved for the following negative gain of amplifier "A":

$$Av=-V_{Y,n,1}/V_{X,n,1}=-1-R/R_S$$

Notice that any undesired small signal that can be modelled by a current source between the drain and source terminals of the matching device is cancelled too. This includes, for instance, 1/f noise and thermal noise of the distributed gate resistance.

The noise factor of the noise cancelling LNA is now determined by the two remaining contributions: the small one of the resistor R[a] and that of amplifier "A". There is however an important difference with respect to the F of the matching stage standalone. The *fixed* contribution of the matching device is now replaced with the *unconstrained* one of the amplifier "A". The latter decreases by increasing the g_mR_S of its input stage, thus enabling noise factors of the LNA well below 2 (i.e. NF below 3dB) for $Z_{IN}=R_S$.

From the above analysis of noise cancelling, two basic characteristics are evident:

- The cancellation is independent on $\alpha(R_S,g_{m,i})$ and so on the quality of the source impedance match. This is because any change in of $g_{m,i}$ affects equally the noise voltage $V_{X,n,i}$ and $V_{Y,n,i}$. In fact, $g_{m,i}$ determines only the amount of noise to be cancelled.
- Noise cancelling depends on the absolute value of the *real* impedance of the source, R_S (e.g.: the impedance seen "looking into" a terminated coax cable).

IV. LNA IC DESIGN

The LNA concept in figure 4 was simulated using models for a 0.25µm MOS process. Amplifier "A" was assumed noiseless. The matching stage provides $Z_{IN}=1/g_{m,i}=R_S=50\Omega$ and $R=300\Omega$ renders a voltage gain of 12.8dB. The NF at 1GHz appears to drop from a maximum of 6dB at Av=0, (i.e. NF of the matching stage standalone) as the gain |Av| increases. A minimum NF of about 0.6dB (i.e. the contribution of R) occurs at |Av|=7. This result is in agreement with $|Av|=1+R/R_S=1+300/50=7$, predicted by the hand analysis. As expected, the cancellation works also at low frequency where the 1/f noise of the MOST is important.

The noise-cancelling amplifier lends itself to the simple circuit implementation shown in figure 5. Amplifier "A" is replaced with the common source stage M2-M3, which renders at the output the voltage at node X time the gain its $Av=-g_{m2}/g_{m3}$ (i.e. node Y grounded). Transistor M3 acts also as a source follower copying the voltage at node Y to the output (i.e. node X grounded).

[a] The noise current of R, $I_{n,R}$, can be split in two correlated sources at the output and the input of the matching stage. The former is cancelled for $Av=-1-R/R_S$ while the other directly adds to the signal source. The latter contribution to F, $I^2_{n,R}/I^2_{n,RS}$, is however small for $R/R_S>>1$, which is the case for a sufficiently large gain.

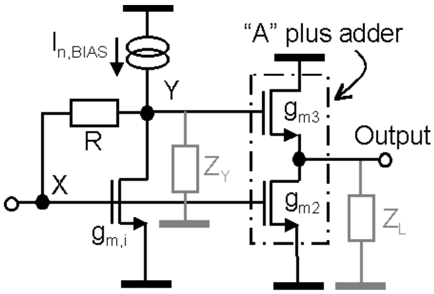


Figure 5: Realisation of the LNA in figure 4 (biasing not shown). The impedance Z_Y and Z_L (grey line) model the output conductance of the MOS and the body effect.

The superposition principle renders the final addition of voltages. Noise cancellation occurs at $g_{m2}/g_{m3} = 1+R/R_S$. The voltage gain $A_{VF}=1-g_{m1}R_S-g_{m2}/g_{m3}$ for $Z_{IN}=R_S$ is then equal to $-2R/R_S$. Note as the noise of the bias current source driving the drain terminal of the matching device is cancelled too. For $|A_{VF}|>>1$, it can be shown that F drops with increasing $g_{m2}R_S$ until it saturates to a minimum of $1-2/A_{VF}$ (i.e. the contribution of R).

Another limitation in F is caused by differences in phase shift between the two noise paths to the output, that limit the amount of achievable noise cancellation.

Compared to the feedback LNA in figure 1g, the noise cancelling offers important advantages:

- Noise cancelling is a feed-forward technique *free* of global feedback, so instability risks are much relaxed.
- The input impedance depends only on the $g_{m,i}$ of the matching MOS. Thus, Z_{IN} is less sensitive to process spread and variable gain at constant $Z_{IN}=R_S$ is more straightforward (e.g.: acting on R or M2-M3).
- Noise cancelling is robust to device parameter variations due to process spread. The reasons are twofold. First, the cancellation equation depends on a *reduced* set of circuit parameters. For instance, the impedance from node Y to ground, Z_Y, and that from the output node to ground, Z_L, do *not* alter the cancellation because they load the two feed-forward paths to the output in the same fashion. Second, for a cancellation error, $\Delta\varepsilon \equiv g_{m2}/g_{m3}-1-R/R_S$, the term in F of the matching device, $NEF\cdot(\Delta\varepsilon/A_{VF})^2$, is only a slowly rising function of $\Delta\varepsilon$ in the range of practical interest.

A wide band noise cancelling LNA was designed in 0.25µm standard CMOS to demonstrate the feasibility of the concept. Since the purpose is to validate the noise cancelling principle, the design focuses on the noise performance. The following requirements common to high-sensitivity receiver applications were used:

- Bandwidth: from MHz to GHz.
- Voltage gain: $A_{VF}=V_{OUT}/V_{IN}>10dB$.
- Impedance matching: $Z_{IN}=R_S=50\Omega$.
- NF: well below 3dB over the signal bandwidth.

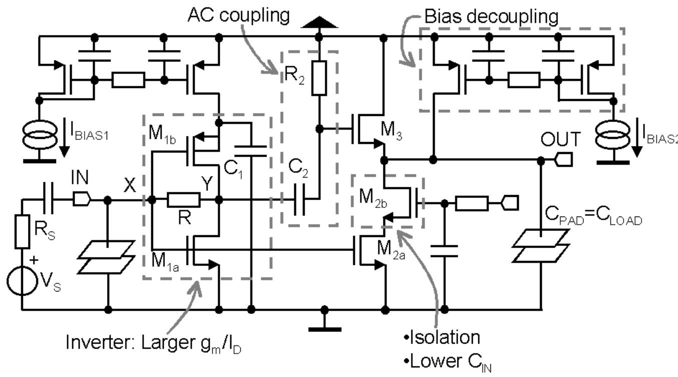


Figure 6: Schematic of the wide band CMOS LNA.

Figure 6 shows the LNA schematic. The matching stage exploits shunt-feedback around a CMOS inverter to provide an input impedance $Z_{IN}=1/(g_{m1a}+g_{m1b})$ and has a larger g_m/I_D. To reduce the sensitivity of gain and input impedance to variations in the supply voltage, the inverter is biased via a current mirror while a MOS capacitor C_1 (i.e. 13pF) grounds the source of M1b. The matching stage is ac-coupled to M3 via the high-pass filter C_2-R_2 (i.e. 0.8pF/95KΩ). The cascode M2b improves the reverse isolation and reduces the input capacitance by decreasing the Miller effect around M2a.The bias current, I_{BIAS2}, allows M3 to conduct only a part of the drain current of M2. This allows to fit within the supply voltage V_{DD}=2.5V without sacrificing NF because the LNA gain is quite large and relatively large voltage headroom is available across the current mirror output. The capacitance of the output bond-pad C_{PAD}=0.2pF is used as load.

Figure 7 shows the result of an simulation experiment on the full circuit design *without* and *with* noise cancellation. This is done by either removing or adding the ac-coupling filter C_2-R_2 (shorting the gate of M3 to the positive supply in case of removal). Without the AC coupling network, the NF is significantly larger (between 5 and 5.5dB) because the LNA acts as a CS amplifier stage with an active input termination of about 50Ω (i.e. the matching stage). As such, NF must exceed $10\cdot Log_{10}(1+NEF)>3dB$ for NEF>1. On the other hand, when the ac-coupling is added, the NF is well below 2.5dB over the full range of frequencies. This can only be explained by the fact that noise cancellation takes place at the output of the LNA.

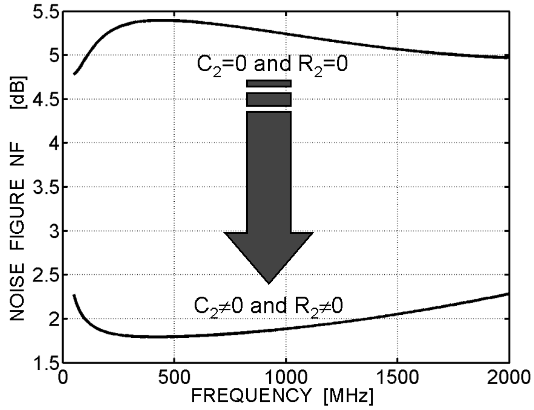


Figure 7: LNA NF w/o ac-coupling filter C_2-R_2.

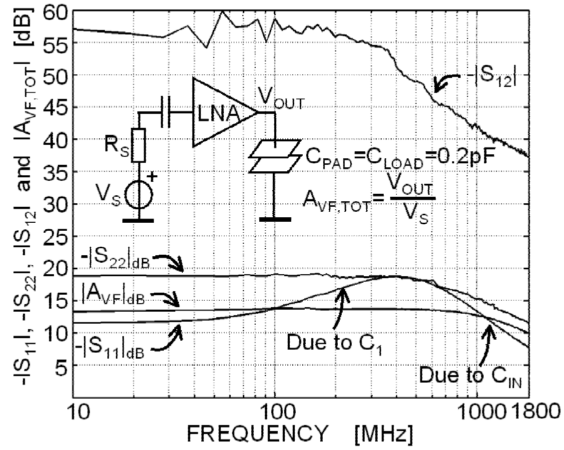


Figure 9: Measured S-parameters and voltage gain.

V. Measurements

Figure 8 shows the chip-photo of the wide band LNA [7]. The LNA S-parameters were measured on-wafer. Figure 9 shows the measured S_{11}, S_{22}, S_{12} and the total voltage gain $A_{VF,TOT}$ from 1 to 1800 MHz.

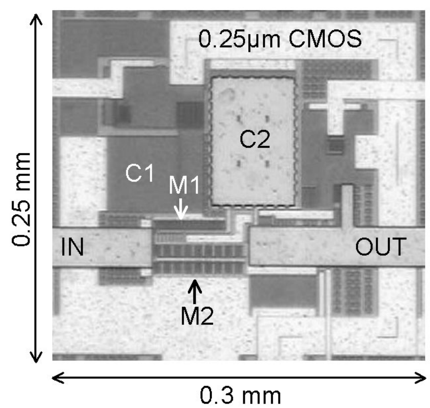
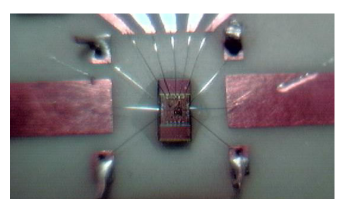


Figure 8: LNA chip-photo.

A flat gain of 13.7dB is found over a –3dB bandwidth between 2MHz and 1600MHz. At 1800 MHz the gain is still 10dB. The reverse isolation $|S_{12}|$, is better than –42dB up to 1GHz and better of -36dB up to 1.8GHz. The input match, $|S_{11}|$ is better than –10dB in 10-1600 MHz and better than –8dB in 10-1800 MHz. At low frequencies, $|S_{11}|$ rises due to the shunt capacitor C_1 in the matching stage. At high frequencies, $|S_{11}|$ drops due to the input capacitance C_{IN}.

Noise figure and distortion properties were measured with the chip-die glued to a low-loss ceramic substrate and connected to 50Ω input/output transmission lines via short bond-wires (about 2mm). Care was taken to minimise the parasitic inductance of the ground return path by connecting several bond-wires in parallel and to de-couple the biasing ports.

Figure 10 shows the printed circuit board (PCB). The NF measurement was also exposed to the effect of the short input bond-wire. The short input bond-wire compensates somewhat the effect of the input capacitance C_{IN}, thereby matching and NF improve at high frequency. A better impedance match is increases the accuracy of the measurement[b] too. The PCB NF was measured with a calibrated HP NF-meter. Losses due to connectors and input/output transmission lines (TLs) were measured separately (on another PCB with the same TLs and connectors) and their effect was de-embedded. Figure 11 shows measured, simulated and calculated NF (taking into account practical differences in phase shift between the 2 noise paths). The measured NF is below 2.4dB over more than one decade (150-2000 MHz) and below 2dB over more than 2 octaves (250-1100 MHz). Next, the agreement with simulation and calculation is good.



a)

b)

Figure 10: PCB for NF and IIPs measurements.

[b] A 3dB-attenuator at the input improves the match of about 6dB.

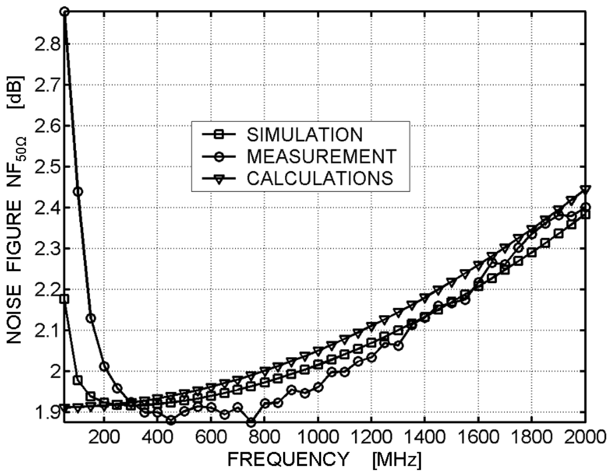


Figure 11: NF vs. frequency.

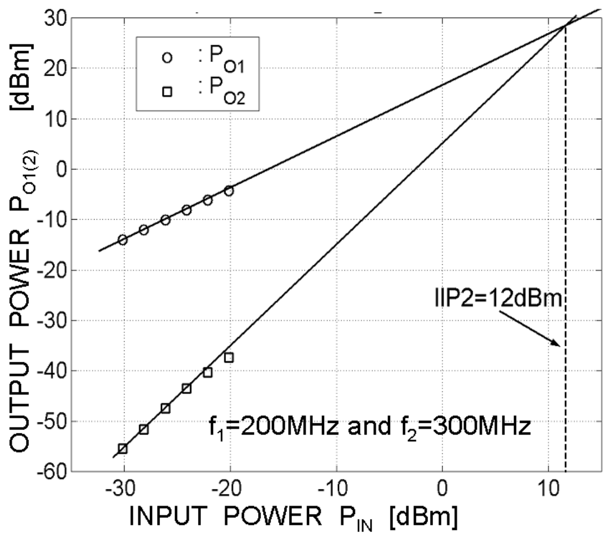


Figure 12: Measured IIP2.

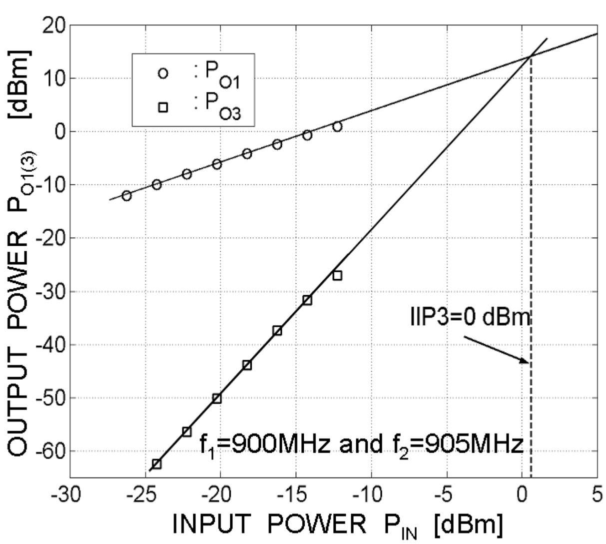


Figure 13: Measured IIP3.

Figures 12 and 13 show the IIP2 and IIP3 for two tones at (200MHz, 300MHz) and (900MHz, 905MHz), respectively. The IIP2 is equal to +12dBm and IIP3 is equal to 0dBm. Table 1 gives a complete summary of the measurements.

PROPERTY	VALUE
$\|A_{VF}=V_{OUT}/V_S\|$	13.7 dB
-3dB BW	2-1600 MHz ($C_{LOAD}=C_{PAD}$=0.2pF)
$\|S_{12}\|$	<-36dB in 10-1800 MHz
$\|S_{11}\|$	<-8dB in 10-1800 MHz
$\|S_{22}\|$	<-12dB in 10-1800 MHz
IIP3 (input ref.)	0 dBm (f_1=900MHz & f_2=905MHz)
IIP2 (input ref.)	12 dBm (f_1=200MHz & f_2=300MHz)
ICP1 (input ref.)	-9 dBm (f_1=900 MHz)
$NF_{50\Omega}$	≤2dB [0.25-1.1 GHz] & ≤2.4dB [0.15-2 GHz]
$I_{DD}@V_{DD}$	14mA@2.5Volt
Area and Technology	0.3x0.25mm² in a 0.25μm CMOS

Table 1: Summary of the measurements.

VI. CONCLUSIONS

In this paper, a novel feed-forward wide-band low-noise technique was presented, which breaks the trade-off between noise figure and $Z_{IN}=R_S$ by cancelling the noise contribution of the matching device at the output of the LNA. This is done without degrading the signal transfer or the quality of the source match. By exploiting this technique it is possible to design wide band LNAs with a very low noise figure over a wide range of frequencies without suffering from instability issues related to the use of negative feedback. Furthermore, the technique is easy to use and robust to device parameter variations. Measurement results on a wide band LNA in 0.25um standard CMOS show NF values below 2.4dB over more than one decade and below 2dB over more than two octaves. Table 3 provides a complete summary of the measurements.

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